ANALOG DEVICES

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APPLICATION NOTE

Microstepping Drive Circuits for Single Supply Systems

by John Wynne

This application note presents a brief review of two-phase permanent magnet stepper motors and outlines some of the issues involved in microstepping. The static position errors introduced by the DACs and by load torque are described and the merits of closed-loop versus open-loop control discussed. Two practical circuits are presented which use positive-only power supplies (+5V and +12V), a situation common in both floppy and low density hard disk drives.

SOME BASICS OF THE TWO-PHASE PM STEPPER MOTOR

In a permanent magnet stepper motor torque is generated by the interaction of two magnetic fields, the stator field and the rotor field. The stator field is generated by current flowing in the phase windings; the rotor field is due to the permanent magnet pole pairs arranged radially on the rotor. A two-phase stepper motor has two phase windings, A and B, separated by 90 electrical degrees, each phase having two possible current directions, positive and negative. Assuming only one phase is turned on at any one time, the stator field can have one of four possible orientations. The rotor field, and hence the rotor, will align itself with respect to the stator field at the one position where the field vectors are in equilibrium.

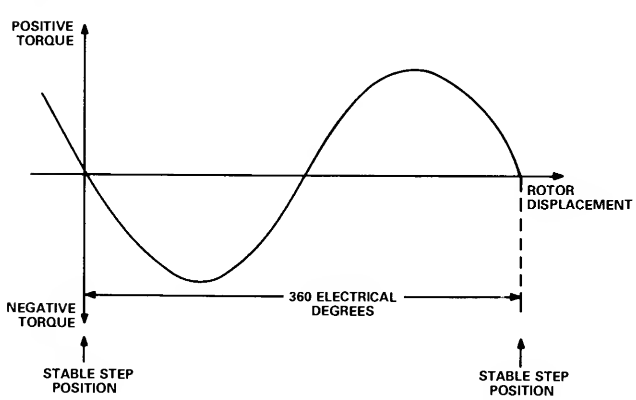

Figure 1. Static Torque vs. Rotor Displacement for an Ideal Stepper Motor

Exactly how the magnetic fields interact to produce torque can easily be seen from the ideal static torque versus rotor displacement curve of Figure 1. This is generated by energizing a single phase and rotating the rotor by hand.

The convention used in Figure 1 is that a movement to the right corresponds to a clockwise rotation of the rotor while a positive torque in the figure represents a clockwise torque developed on the rotor. The torque output can be seen to be a sinusoidal function of the electrical angle ϕ between the field vectors or:

$$T = -K\,I\,\mathrm{Sin}\phi \tag{1}$$

where K is a motor constant and I is the phase current. The maximum torque T_{max} exerted on the rotor occurs when the angle between the rotor and stator field vectors equals 90 electrical degrees. With maximum rated current I_R flowing, this maximum torque is known as the peak static torque or simply the holding torque, i.e., $T_{HMAX} = KI_R$ which is a constant for a given motor. The minus sign in the torque expression indicates that the position of stable equilibrium occurs on the negative slope of the torque curve. If the rotor is moved in a clockwise direction, a counterclockwise torque is developed to return the rotor to its stable or step position. A similar situation exists if the rotor is moved in a counterclockwise direction. Thus a constant phase excitation provides one stable rotor position every 360 electrical degrees. If the phase excitation is reversed, a new stable position will be established midway between these original positions, i.e, at 180 degrees. In a two-phase motor, the two-phase windings are physically separated by 90 electrical degrees, hence stable rotor positions will occur at every 90 electrical degree rotation of the stator field. A two-phase PM stepper will therefore move four full steps for every 360 electrical degree rotation of the stator field vector.

The number of pole pairs N determines the relationship between a stator field vector shift of α electrical degrees

and the resulting rotor movement of θ mechanical degrees;

$$\alpha = N\theta \qquad (2)$$

The full step mechanical angle is therefore

$$\theta_{FS}=\pi/2N \qquad (3)$$

For instance a rotor magnetized with 25 pole pairs will have a mechanical step angle of 3.6°. Figure 2 shows the complete static torque versus rotor position curves for a two-phase permanent magnet stepper motor with each phase excited in turn. Stable equilibrium positions for the rotor exist at the full step positions, $\pi/2N$, π/N, $3\pi/2N$ and $2\pi/N$. A full step rotation of the rotor results from a shift in the stator field vector of 90 electrical degrees. The basic sequence for a two-phase motor of 4 full steps for one full revolution (360 electrical degrees) of the stator field vector is obvious from Figure 2.

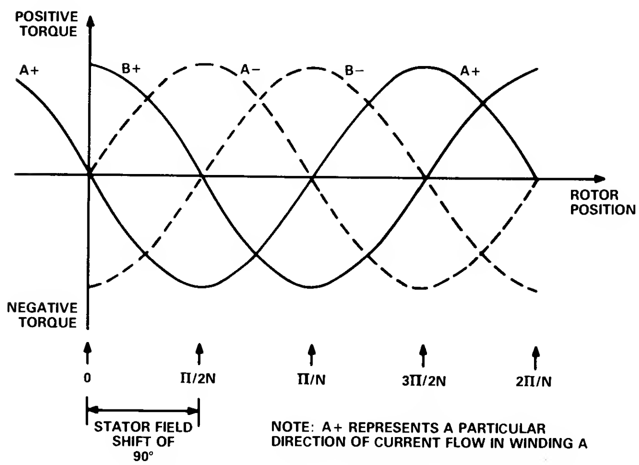

Figure 2. Static Torque vs. Rotor Position Curves for a Two-Phase, PM Stepper Motor

SOME MICROSTEPPING BASICS

In Figure 2 at rotor position 0, phase A current is positive and equal to its maximum rated value I_R; phase B current is zero. At rotor position $\pi/2N$, phase B current is positive and equal to its maximum rated value I_R while phase A current is zero. By correctly choosing and controlling the ratio of phase A current to phase B current it is possible to generate a stable equilibrium position anywhere between the stable full step positions. Figure 3 is a vector representation of the two-phase currents, I_A and I_B, where only the first step from the sequence of four in Figure 2 is represented.

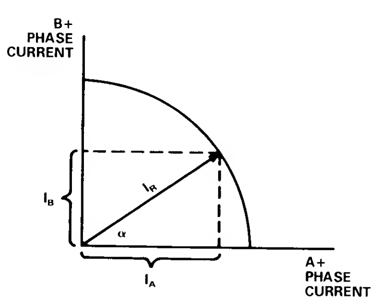

Figure 3. Vector Representation of Phase Currents

From Figure 3 the angular position of the stator field vector, and hence the rotor position, is proportional to the phase currents:

$$\alpha = \text{Arc tan } (I_B/I_A) \qquad (4)$$

In a practical circuit dual digital-to-analog converters (DACs) set the required phase current magnitudes while pulse width modulator (PWM) controllers ensure the set levels are met and maintained in the windings. The maximum number of current ratios or microsteps possible is determined by the resolution n of the DACs, i.e., maximum number of microsteps $= 2^n$. In practise the number of microsteps used is usually an integer power of 2, i.e,. 8,16,32, etc. There are advantages to using DACs whose resolution exceeds the minimum resolution necessary to generate the required number of microsteps.

Not alone is it necessary to choose the correct ratios of phase currents for each new microstep, but it is also necessary to ensure that the holding torque due to the combined phases remains constant and equal to the holding torque obtained with full rated current in a single phase, i.e., from Figure 3

$$\sqrt{I_A^2+I_B^2}=I_R=\text{Constant} \qquad (5)$$

If this condition is not met, it is possible, with a purely friction load, to have a different position error at each microstep. This is discussed later.

The torque output generated by the two motor phases is:

$$T_A = -KI_A \text{ Sin}\phi \qquad (6)$$

$$T_B = -KI_B \text{ Sin } (\phi - \pi/2)$$

$$\text{or } T_B = KI_B \text{ COS } \phi \qquad (7)$$

To move the rotor by θ mechanical degrees (or the stator field vector by $N\theta$ electrical degrees) it is necessary to vary the phase currents in a cosinusoidal fashion:

$$I_A = I_R \text{ Cos}N\theta \qquad (8)$$

$$I_B = I_R \text{ Sin}N\theta \qquad (9)$$

In an ideal motor both torque contributions can be added to provide the total torque:

$$T_{TOT} = T_A + T_B$$

$$= -KI_R \text{ Cos}N\theta \text{ Sin}\phi + KI_R \text{ Sin}N\theta \text{ Cos}\phi$$

$$= -KI_R (\text{Sin}\phi \text{ Cos}N\theta - \text{Cos}\phi \text{ Sin}N\theta)$$

$$T_{TOT} = -KI_R\text{Sin } (\phi - N\theta) \qquad (10)$$

If the stable torque equilibrium position is moved by $N\theta$ electrical degrees, the rotor will follow it by moving θ mechanical degrees to where the torque exerted on the rotor is zero. Varying the phase currents in a sine-cosine relationship provides the required microstep movement of the stator field vector and ensures a constant stator field vector magnitude (or holding torque) at each microstep position.

STATIC POSITION ERRORS

Microstepping accuracy depends on how closely equations 6–10 can be realized in practise with a real motor. The perfectly sinusoidal static torque vs. rotor displacement curve exhibited by an ideal motor can be upset by a number of factors such as the presence of detent torque, a nonlinear relationship between torque and energizing current, by nonidentical static torque vs. angular displacement curves for Phase A and Phase B, etc. It is worthwhile analyzing the positional errors which can be caused by the PWM controllers and the DACs.

PHASE GAIN ERRORS

One of the most common causes of inaccuracies in microstepping is unequal maximum torque contributions from both phases. This can be due to the motor itself, to mismatch in the PWM controllers (for instance, due to different values of sense resistors) or to mismatch in the DAC outputs (for instance, different gain errors). The torque expressions (Equations 6 and 7) can be rewritten to include a phase gain error:

$$T_A = -KI_A \operatorname{Sin}\phi \tag{11}$$

$$T_B = (1+M)\, KI_B \operatorname{Cos}\phi \tag{12}$$

where $I_A = I_R \operatorname{Cos}N\theta$ (8)

$I_B = I_R \operatorname{Sin}N\theta$ (9)

and M is the mismatch between phases. The stable rotor position occurs when $T_A + T_B = 0$:

$$-KI_R \operatorname{Cos}N\theta \operatorname{Sin}\phi + (1+M)\, KI_R \operatorname{Sin}N\theta \operatorname{Cos}\phi = 0 \tag{13}$$

$$\text{or } \tan\phi = (1+M)\tan N\theta \tag{14}$$

$$\text{or } \phi = \operatorname{Arc}\tan[\,(1+M)\tan N\theta] \tag{15}$$

For an ideal motor at its stable rotor position $\phi = N\theta$. The difference between ϕ and $N\theta$ at the stable rotor position indicates positional error:

$$\phi - N\theta = \operatorname{Arc}\tan[(1+M)\tan N\theta\,] - N\theta \tag{16}$$

Equation 16 is plotted in Figure 4 for two phase gain error values of 1% (M = 0.01) and 5% (M = 0.05). The horizontal axis of Figure 4 is the rotor position in electrical degrees. Maximum error occurs at the half-step position. For example, in a system with sixteen microsteps per full step, a phase gain mismatch of 5% results in a microstep position error of nearly 25%.

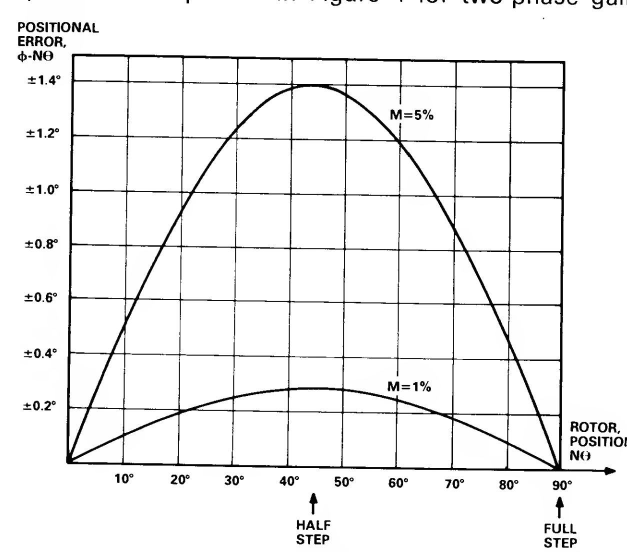

Figure 4. Microstepping Positional Error Due to Phase Gain Errors

ERRORS DUE TO DAC RESOLUTION AND ACCURACY

From before, the angular position of the rotor is proportional to the ratio of phase currents

$$\alpha = \operatorname{Arc}\tan\,(I_B/I_A) \tag{4}$$

The locus of ideal phase current ratios required to microstep the rotor between two full-step positions lies along the quarter circle in Figure 3. However, due to the finite resolution of the DACs in a practical system, it is not possible to choose the ideal current ratios required at each microstep to exactly track the circular locus. Consequently, the DAC finite resolution introduces an angular error in the different positions of the current vector. The angular error at each microstep can be expressed as

$$\delta_\alpha = \operatorname{Arc}\tan\,(I_B/I_A) - \alpha_{MICROSTEP} \tag{17}$$

where $\alpha_{MICROSTEP}$ is the ideal microstep size in any given system. For example, Figure 5 compares the percentage angular errors generated by two different resolution DACs (6-bits and 8-bits) for a system with 8 microsteps per full step (i.e., $\alpha_{MICROSTEP} = 90°/8 = 11.25°$). Figure 5a shows that the limited resolution available with a 6-bit DAC can contribute almost 5% angular error to the microstep position; whereas with an 8-bit DAC, Figure 5b, the errors are approximately an order of magnitude less.

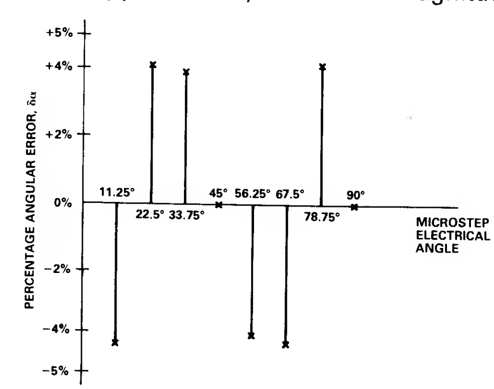

Figure 5a. Percentage Angular Error for 8-Microstep System Using Ideal 6-Bit DACs

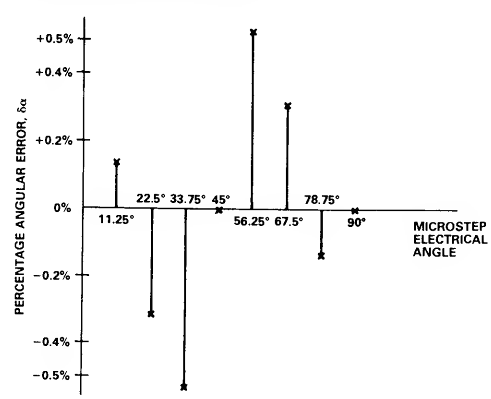

Figure 5b. Percentage Angular Error for 8-Microstep System Using Ideal 8-Bit DACs

The graphs of Figure 5 assume the DACs have no integral linearity errors. If the DACs are assumed to be accurate to their resolution level, i.e., the 6-bit DACs are 6 bits accurate (±0.8%) and the 8-bit DACs are 8 bits accurate (±0.2%), the errors increase over the previous case. Figure 6 repeats the comparsion using such DACs. Maximum angular error using a 6-bit accurate DAC can now be up to 10%.

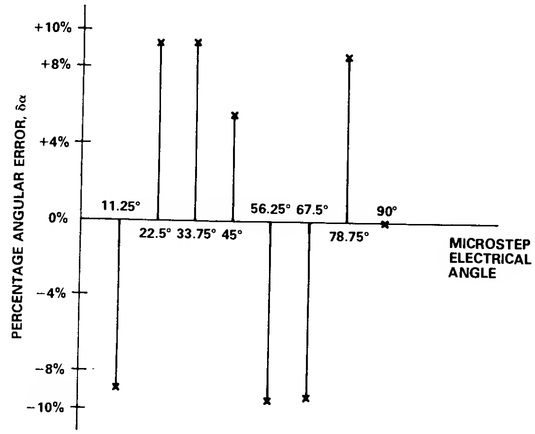

Figure 6a. Percentage Angular Error for 8-Microstep System Using 6-Bit Accurate DACs

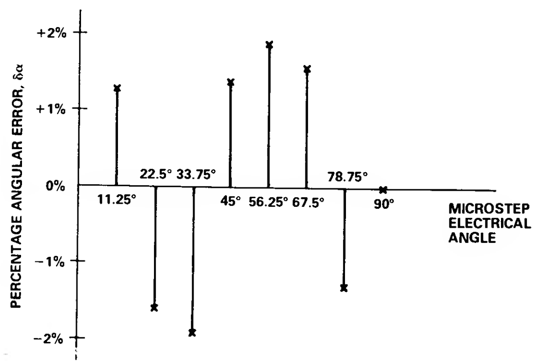

Figure 6b. Percentage Angular Error for 8-Microstep System Using 8-Bit Accurate DACs

EFFECT OF LOAD TORQUE

One of the most significant error sources in a microstepper positioning application is load torque T_L. The effect of load torque can be seen from the typical static torque vs. rotor position curve in Figure 7.

With no load, the rotor is at rest at position 0. Applying a clockwise load torque T_L to the rotor will move the rotor to a new rest position, a position at which the applied clockwise torque T_L is balanced by the anticlockwise torque developed by the motor. The rotor displacement between the two rest positions is the step position error θe. This can be expressed as

$$\theta e = \frac{\text{Arc Sin}\,(-T_L/T_{HMAX})}{N} \qquad (18)$$

For a given load torque the step position error is inversely proportional to the number of pole pairs on the rotor and to the maximum holding torque available. Increasing the number of pole pairs decreases the full step size thereby increasing the slope of the static torque

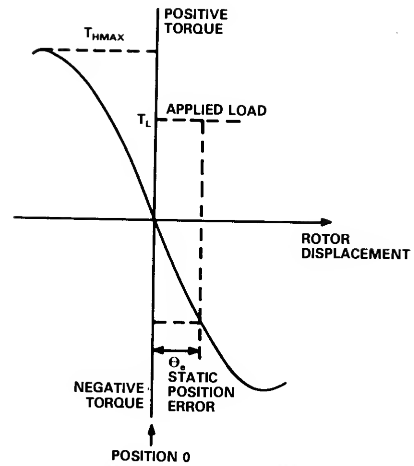

Figure 7. Static Position Error Due to Load Torque

vs. rotor displacement curve over that obtainable with a bigger step size motor. For two motors with equal load torque and holding torque but with different step sizes the steeper slope will result in smaller step position errors. However, not alone do motor costs obviously increase with smaller step size, but so too does the static torque vs. displacement curve depart from the ideal sinusoidal relationship. Specialist motors are available from manufacturers which have been deliberately optimized for microstepping during manufacture.

The other option for reducing step position error indicated by Equation 18, that of increasing the holding torque (by increasing the phase currents), is not available when microstepping since the cosinusoidal phase currents ensure the exact opposite occurs, i.e., that the holding torque remains constant at each microstep. Thus, when microstepping, the maximum holding torque is deliberately restricted to equal the one-phase-on torque, not the higher torque which normally would be available with two-phase-on operation.

CLOSED-LOOP VS. OPEN-LOOP CONTROL

Microstepping improves the positioning resolution possible in any control application. However, the positional accuracy can be significantly worse than that offered by the original full-step accuracy specification due primarily to load torque effects. To ensure that the increased resolution is useable it is necessary to change from a normally open-loop system, where the rotor is simply directed to move so many microsteps, to a closed-loop system where the actual movement of the rotor is monitored and any residual error between demanded position and actual position can be detected and subsequently corrected. The correction is accomplished by adjusting the ratio of the phase currents to pull the rotor into the demanded position. For two systems with the same number of microsteps per full step (e.g., 16, 32, etc.) but with one using 6-bit DACs and the other 8-bit DACs, the higher resolution available from the 8-bit DACs offers a greater number of current ratios to be selected allowing a more accurate final rotor position to be achieved. Thus, fine positioning with high accuracy can be achieved with microstepping but only in a closed-loop system.

PRACTICAL CIRCUITS

Two typical microstepping drive circuits are presented in this application note. Both use dual 8-bit DACs to control the phase currents and a single 8-bit ADC for closed-loop position control. The first circuit, Figure 9, uses an AD7628 dual DAC and an AD7820 ADC. The AD7628 is identical to the industry standard AD7528 but offers TTL compatibility (and is fully specified) at a +12V power supply. It is beyond the scope of this application note to deal in any detail with the types of position/velocity transducers which are used to monitor movement of the head assembly in disk drives. The AD7820 half-flash 8-bit ADC is widely used for converting these transducer output signals and it is shown in block diagram form in Figure 9. For example, a typical transducer might be an incremental optical encoder with quadrature triangular output waveforms although stepper motors are now becoming available (such as the Portescap P750) which include, in addition to the phase windings, integral velocity windings which can be used directly to obtain rotor velocity and hence position. The second circuit, Figure 11, uses a recently released device from Analog Devices, the AD7669, which contains both a dual 8-bit DAC and an 8-bit ADC.

Typical DAC output voltage waveforms and associated direction control information for clockwise rotation are shown in Figure 8.

The "sine/cosine" data is usually held in look-up tables in the controller. The rate at which the data is loaded to the DACs obviously determines the microstepping rate. The controller also supplies the direction control information required by the H-bridge power driver. The output driver used in both circuits is the UDN2998W, a dual H-bridge driver available from Sprague Electric. It is packaged in a 12-pin single-in-line power tab package

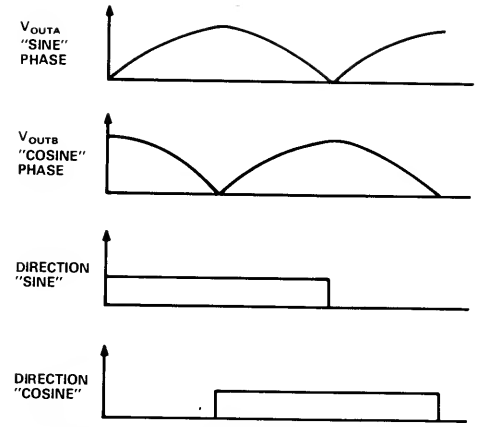

Figure 8. Typical DAC Output Voltages for Microstepping. Direction Signals Are for CW Rotation with UDN-2998W.

for high current capabilities (+2A continuous). In practice, it was found necessary to ground the power tab to a quiet ground in order to avoid parasitic oscillations.

The circuits of Figures 9 and 11 also differ from each other in that one is based on a fixed frequency PWM technique while the other uses a frequency modulation PWM approach. The fixed frequency approach allows a synchronized chopping frequency which can help to reduce noise on the power supply lines and ground return paths. The frequency modulation approach provides a fixed-ripple current characteristic leading to a higher efficiency drive.

FIXED FREQUENCY PWM DRIVE

The complete circuit is shown in Figure 9. The NE555 timer produces short (2.5µs) low level pulses at a repetition rate of approximately 45kHz. This is high enough to avoid audible noise and low enough to avoid unnecessary switching losses. These negative pulses reset

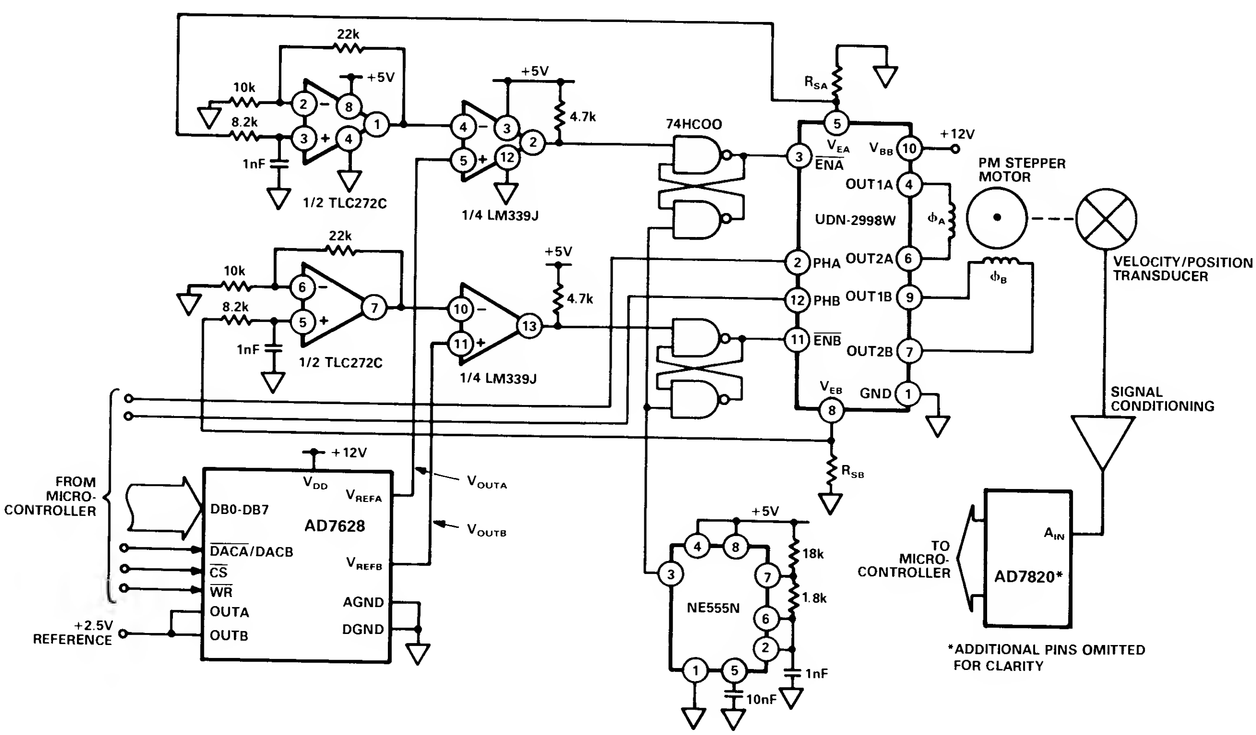

Figure 9. Fixed Frequency PWM Drive

the flip-flops and enable the phase currents via the UDN2998W dual H-bridge driver. The actual value of the phase currents are monitored by means of the sense resistors R_{S1} and R_{S2} between the emitters of the output sink drivers and ground. The voltage signals across the sense resistors — representing the phase currents — are filtered by the 8.2kΩ/lnF combination to eliminate noise spikes, amplified by X3.2 across I_{C1} a/b and compared with the DAC control voltages by I_{C2} a/b. For either channel, when the amplified sense voltage exceeds the DAC control voltage, the comparator output is driven low setting the flip-flop and turning off the source driver. The phase current now decays slowly, the recirculation path being through the on-chip ground-clamp diode, phase winding and output sink transistor. The current will continue to fall until the next pulse resets the flip-flop and the sequence is repeated.

The lower trace of Figure 10 shows a typical output voltage waveform from one of the DACs. The waveform covers two full steps with eight microsteps per full step. The top trace is the amplified sense voltage at the output of the corresponding sense amplifier. The vertical axis is 1V/division. The horizontal axis is uncalibrated for photographic purposes, but the time for eight microsteps is 12.8ms. The motor used is a Portescap P530 two-phase stepper motor with a full step size of 3.6°. It is intended for 0.45° microstepping −eight microsteps per full step — resulting in a total of 800 microsteps per revolution. With a reference voltage of +2.5V on the DACs and a sense resistor of 0.5Ω, the peak current is (2.5/3.2)/0.5 or 1.56A which is equal to the nominal rated current for one-phase on operation.

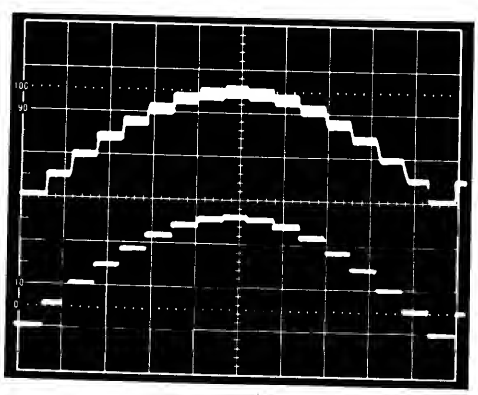

Figure 10. Circuit Waveforms for Fixed Frequency PWM Circuit of Figure 9

VOLTAGE-MODE DACs

The AD7628 DACs are operated in what is known as a voltage-mode or voltage-switching configuration. The reference voltage is applied to OUTA and OUTB terminals of the AD7628, and the output is taken from the V_{REFA} and V_{REFB} terminals. A positive reference voltage gives a positive output voltage — hence the output signals are voltages (not currents) at a constant source impedance equal to the ladder resistance and can be expressed as

$$V_{OUTA} = D_A \cdot V_{REF} \tag{19}$$

$$V_{OUTB} = D_B \cdot V_{REF} \tag{20}$$

where $D_A = N_A/256$ and $D_B = N_B/256$, with N_A and N_B in hexadecimal format being the codes supplied to DAC A and DAC B respectively. This mode of operation is virtually free of gain error, hence DAC to DAC output voltage matching will be better than 0.4%. Close matching is important in open-loop micropositioning applications since inequalities between the maximum values of phase currents will cause positioning errors.

The circuit of Figure 9 responds to DAC output voltages extending from V_{REF} down to and including 0V. An operating range which includes 0V is important in applications such as disk drive read/write head movement which require very fine positioning indeed. Such systems can have 32, 64 or more microsteps per full step producing very small voltage steps around 0V. These occur when the rotor position is close to a full step position. Also at this position the phase with the least current in it has the biggest impact on the positioning error.

FREQUENCY MODULATION PWM DRIVE

By removing the flip-flop from Figure 9 which synchronizes the phase-on times with a master clock frequency, the resultant circuit will free-run at a rate determined by the response of the loop components and the motor parameters. This circuit is shown in Figure 11.

For any one phase assume that the comparator output is high indicating that the phase current is less than it should be. The high on the comparator output produces a low on the enable input of the H-bridge and turns on the phase current. The phase current is monitored by the sense resistor, filtered and amplified before being compared with the control voltage. When the amplified sense voltage exceeds the DAC control voltage, the phase current is turned off. The comparator output remains low until the decaying current falls sufficiently to cause the comparator output to go high again, restarting the cycle. Although both phases have separate loops, in practice the chopping frequencies lock together avoiding any beat phenomena. With the components shown in Figure 11, and again using the Portescap P530 stepper motor, the chopping frequency is approximately 25kHz. It should be possible to reduce the component count of Figure 11 by replacing the dual op amp and quad comparator (half of which is used) by one of the modern quad op amps designed for single supply operation. Due to the slower response time of the op amps when used as comparators the chopping frequency will drop but the circuit should still operate due to its regenerative nature.

It is also very simple to add an extra control signal to allow the phase currents to be turned off independent of the controller. By replacing the inverter gates shown in Figure 11 with NAND gates, the spare input on each NAND gate can now act as a control, high for normal operations; low for emergency stop.

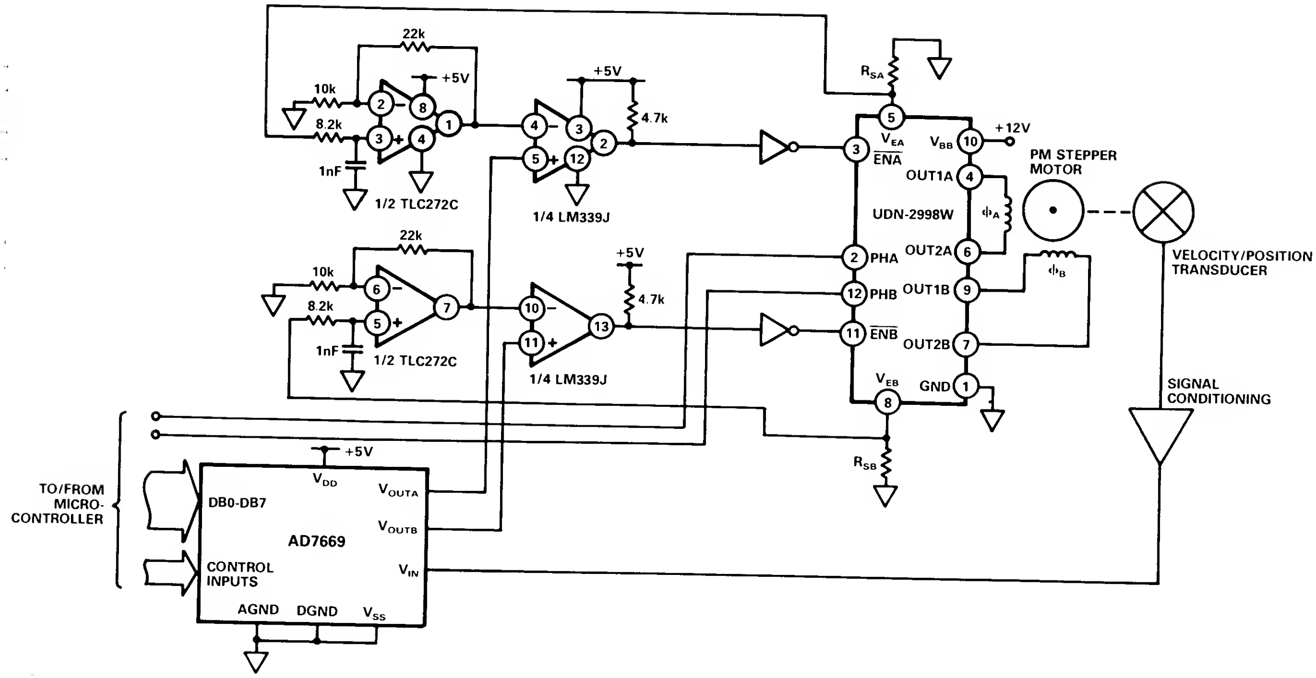

Figure 11. Frequency Modulation PWM Drive

The top trace in Figure 12 shows the amplified sense voltage at the output of one of the sense amplifiers. The lower waveform shows the DAC control voltage for that phase. To allow a direct comparsion with the previous waveform of Figure 10 the vertical axis of Figure 8 is again 1V/division with an identical microstepping rate of 1 microstep per 1.6mS or 8 microsteps per 12.8mS. The fixed-ripple current characteristic of the drive is very obvious.

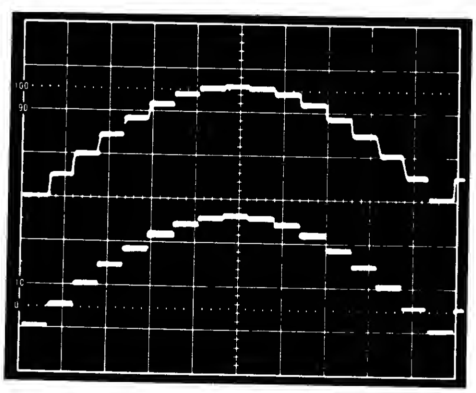

Figure 12. Circuit Waveforms for Frequency Modulation PWM Circuit of Figure 11

VOLTAGE OUTPUT DACs

The two DACs used in this circuit are 8-bit voltage DACs contained on the monolithic AD7669 die. An on-chip bandgap voltage reference of +1.23 (not available to users) drives both DAC inputs. The output buffer amplifiers of each DAC are configured with internal X2 gains and provide unipolar 0 to +2.5V outputs. DAC-to-DAC full-scale error matching is better than ±0.4%. As mentioned previously, the AD7669 also contains an 8-bit ADC. The ADC input voltage range matches the DACs' output voltage range of 0 to +2.5V with a typical DAC-to-ADC gain match of 1%.

MICROSTEPPING RATE

Outside of disk drive applications where microstepping is used to position the read/write head, its primary attraction is the extremely smooth single step response which can be achieved by replacing a single excitation change with a number of smaller amplitude excitation changes. Such applications are inherently low speed – below 100 rpm – since at higher speeds the motor inertia makes the rotor rotate smoothly regardless of whether the drive is microstep, full step or half step. In head positioning applications in disk drives motor speed will depend on whether the drive is in the seek or track mode. In the seek mode where the head is moving some distance across the disk surface, highest possible speeds are required to meet the velocity profile. In the track mode where the head is tracking a selected data track, the microstepping rate is lower. For the fixed frequency PWM circuit of Figure 9, the maximum microstep rate before any individual microstep disappeared was found to be 1400 microsteps/second. With 0.45° microsteps this rate translates into 105 rpm. The rate can be much increased above this before the motor stalls, but the sense voltage waveforms increasingly depart from the DAC control waveforms. With 0.225° microsteps (16 per full step) the maximum microstep rate which retains all of the microsteps in the sense waveforms was found to be 1850 microsteps/sec or approximately 70 rpm. Similar results were found for the frequency modulation PWM circuit of Figure 11. With 0.45° microsteps, maximum microstep rate retaining all eight levels was found to be 1330 microsteps/sec or 100 rpm. With 0.225° microsteps the rate was 1660 microsteps/sec or approximately 62 rpm.

NEW PWM MOTOR DRIVER AVAILABLE

Sprague Electric has recently released the UDN2917, a new dual full-bridge stepper motor driver suitable for microstepping applications. Output current in both windings of a bipolar stepper motor is sensed and controlled independently in each bridge by an external sense resistor, internal comparator and monostable multivibrator. When the bridge is turned ON, current increases in the motor winding, and it is sensed by the external sense resistor until the sense voltage reaches the level set at the comparator's input. The comparator then triggers the monostable which turns OFF the source driver of the bridge. After turn-off, the motor current decays, circulating through the ground clamp diode and sink transistor. The source driver's OFF time is determined by the monostable's external RC timing components. When the source driver is re-enabled, the winding current (the sense voltage) is again allowed to rise to the comparator's threshold. This cycle repeats itself, maintaining the average motor winding current at the desired level. This is a frequency modulation PWM drive.

The outputs are capable of driving motors at 1.5A and 45V. Maximum combined (source and sink) output saturation voltage is 2.6V at 1A and 3.1V at 1.5A. A lower power version, the 2916, is also available. The 2917 is available in two packages for applications versatility. One option is a 28-pin DIP with the two center pins on each side bonded together (commonly known as a batwing) and connected to the die attach pad to form a heat tab. For surface mount applications, a 44-pin PLCC is available with two sides of this quad package formed into heat sinking tabs connected to the die attach pad.